

带 LC 滤波器的永磁同步电机多步模型预测 电流控制

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Multi-Step Model Predictive Current Control of Permanent Magnet Synchronous Motor with LC Filter

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Abstract: [Objective] In order to improve the control performance of model predictive current control (MPCC) for permanent magnet synchronous motor (PMSM), an LC filter was added between the PMSM and the inverter. [Method] A third-order multi-step MPCC system with LC filter was established. Predictive control was applied to the inverter current, capacitor voltage, and motor current, with simulation comparisons made against a first-order multi-step MPCC and field orientation control (FOC). [Results] Comparison results showed that under the same switching frequency and number of prediction steps, the third-order MPCC significantly reduced torque ripple and current ripple in the PMSM compared to the first-order MPCC, while also achieving lower total harmonic distortion (THD) in the current. Additionally, as the number of prediction steps increased in the third-order MPCC, both current THD, current ripple, and torque ripple progressively decreased. [Conclusion] As the number of prediction steps increases, the control performance of the third-order MPCC system gradually improves. The third-order MPCC outperforms the first-order MPCC. At higher switching frequencies, the third-order MPCC with 4-5 prediction steps delivers better control performance than FOC.

Key words: permanent magnet synchronous motor; model predictive current control; multi-step prediction; LC filter

摘要: [目的]为提升永磁同步电机(PMSM)模型预测

基金项目:长安大学 2024 年研究生科研创新实践项目
(300103724044)

Scientific Innovation Practice Project of Postgraduates of
Chang'an University (300103724044)

电流控制(MPCC)的控制性能,在 PMSM 与逆变器之间增加 LC 滤波器。【方法】建立带 LC 滤波器的三阶多步 MPCC 系统,对逆变器电流、电容电压及电机电流进行预测控制,并与一阶多步 MPCC 和磁场定向控制(FOC)进行仿真对比。【结果】仿真结果表明:在相同开关频率和预测步数下,相较于一阶 MPCC,三阶 MPCC 下 PMSM 的转矩脉动和电流脉动明显减小,电流总谐波失真(THD)更低;对于三阶 MPCC,随着预测步数的增加,电流 THD、电流脉动和转矩脉动逐步减小。【结论】随着预测步数的增加,三阶 MPCC 系统的控制性能逐步提升;三阶 MPCC 的控制性能优于一阶 MPCC;在较高开关频率处,四步至五步的三阶 MPCC 的控制性能优于 FOC。

关键词: 永磁同步电机;模型预测电流控制;多步预测;LC 滤波器

0 引言

模型预测电流控制(Model Predictive Current Control, MPCC)结构简单,控制灵活,无需脉宽调制,近年来成为永磁同步电机(Permanent Magnet Synchronous Motor, PMSM)领域的研究热点^[1-9]。由于备选电压矢量有限,电流和转矩脉动较大,MPCC 的控制性能不及传统磁场定向控制(Field Orientation Control, FOC)。为了提升有限集模型预测控制的控制性能,在电机驱动系统中增加滤波器,将控制系统从一阶系统变为高阶系统,从而优化控制性能。文献[10]针对电压源逆变器建立 LC 滤波器的数学模型,根据其系统图求得滤波器的传递函数,绘制相应的 Bode 图,证明 LC 滤波器的稳定性和控制效果。文献[11]针对 LC 滤

滤波器的响应频率提出一种设计方法,并验证了该方法具有较优的控制效果。文献[12-15]对电压源逆变器驱动的电机电控系统采用 LC 滤波器,使用模型预测控制,将滤波器电流误差和电压误差以及电机电流误差作为成本函数,发现使用 LC 滤波器的电机驱动系统电流纹波大大降低。同时,将预测步长由传统的单步预测变为多步预测也可提升控制性能^[16-22]。但目前多步模型预测控制大多针对传统一阶系统,未对带滤波器的高阶系统展开研究。

本文在 PMSM 传统一阶 MPCC 的基础上,在逆变器和电机之间增加 LC 滤波器,建立了带 LC 滤波器的三阶多步模型预测电流控制系统。对逆变器电流、电容电压及电机电流进行预测控制,设计包括逆变器电流误差、电容电压误差、电机电流误差以及开关次数的成本函数,对比研究三阶多步 MPCC、一阶多步 MPCC 及 FOC 的控制性能。

1 三阶多步模型预测电流控制

1.1 预测模型

带 LC 滤波器的 PMSM 示意图如图 1 所示。图 1 中: u_{inv} 和 i_{inv} 分别为逆变器端输出的电压和电流; u_s 和 i_s 分别为负载端输入的电压和电流; u_c 为电容两端电压。

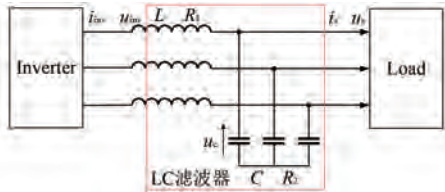


图 1 交流三相 LC 滤波器电路

Fig. 1 AC three-phase LC filter circuit

将 LC 滤波器和 PMSM 作为整个控制对象,输入为逆变器的输出电压,输出为 PMSM 电流。LC 滤波器的电容和电感除了本身感抗和阻抗特性外,还存在等效的损耗电阻,在建立预测模型时将等效损耗内阻考虑在其中,同时将 PMSM 当作感性带阻负载。电机的定子电感和定子绕组分别为 L_s 和 R_s ;电感的感抗和其内部等效电阻分别为 L 和 R_1 ;电容的容抗和其内部的等效电阻分别为 C 和 R_2 。

将三相交流系统转换为单相系统,同时考虑电感等效损耗内阻 R_1 和电容等效损耗内阻 R_2 ,

得到等效电路如图 2 所示。

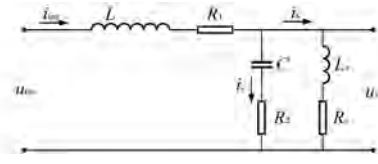


图 2 交流单相 LC 滤波器电路

Fig. 2 AC single-phase LC filter circuit

根据基尔霍夫定律,LC 滤波器的数学模型如式(1)~式(4)所示:

$$u_{inv} = L \frac{di_{inv}}{dt} + R_1 i_{inv} + u_c + R_2 i_c \quad (1)$$

$$C \frac{du_c}{dt} = i_c \quad (2)$$

$$u_s = u_c + R_2 i_c \quad (3)$$

$$i_c = i_{inv} - i_s \quad (4)$$

在 d - q 旋转坐标系下,逆变器电流和电容电压如式(5)和式(6)所示:

$$\frac{di_{inv-dq}}{dt} = \frac{1}{L} [u_{inv-dq} - R_1 i_{inv-dq} - R_2 (i_{inv-dq} - i_{s-dq})] \quad (5)$$

$$\frac{du_{c-dq}}{dt} = \frac{1}{C} (i_{inv-dq} - i_{s-dq}) \quad (6)$$

采用一阶前向欧拉离散法对逆变器电流和电容电压的微分项进行离散化,得到逆变器电流 i_{inv} 、电容电压 u_c 的离散表达式分别如式(7)和式(8)所示:

$$\begin{cases} i_{inv-d}(k+1) = \frac{T_s}{L} u_{inv-d}(k) + \left[1 - \frac{T_s}{L}(R_1 + R_2)\right] \cdot \\ i_{inv-d}(k) - \frac{T_s}{L} u_{c-d}(k) + \frac{T_s R_2}{L} i_{s-d}(k) \\ i_{inv-q}(k+1) = \frac{T_s}{L} u_{inv-q}(k) + \left[1 - \frac{T_s}{L}(R_1 + R_2)\right] \cdot \\ i_{inv-q}(k) - \frac{T_s}{L} u_{c-q}(k) + \frac{T_s R_2}{L} i_{s-q}(k) \end{cases} \quad (7)$$

$$\begin{cases} u_{c-d}(k+1) = u_{c-d}(k) + \frac{T_s}{C} i_{inv-d}(k) - \frac{T_s}{C} i_{s-d}(k) \\ u_{c-q}(k+1) = u_{c-q}(k) + \frac{T_s}{C} i_{inv-q}(k) - \frac{T_s}{C} i_{s-q}(k) \end{cases} \quad (8)$$

式中: T_s 为采样周期。

PMSM 在 d - q 旋转坐标系下定子电流 i_s 的离散表达式如式(9)所示:

$$\begin{cases} i_{s-d}(k+1) = \left(1 - \frac{R_s T_s}{L_d}\right) i_{s-d}(k) + \frac{T_s L_q}{L_d} \omega_e(k) i_{s-q}(k) + \frac{T_s}{L_d} u_{s-d}(k) \\ i_{s-q}(k+1) = \left(1 - \frac{R_s T_s}{L_q}\right) i_{s-q}(k) - \frac{T_s L_d}{L_q} \omega_e(k) i_{s-d}(k) - \frac{T_s \psi_f}{L_q} \omega_e(k) + \frac{T_s}{L_q} u_{s-q}(k) \end{cases} \quad (9)$$

式中: ω_e 为转子电角速度; ψ_f 为转子永磁体磁链。

带 LC 滤波器的 PMSM 模型预测控制系统将逆变器输出电压当作控制系统的输入, 采用 $u_s = u_c + R_2 i_c$ 将 d 、 q 轴定子电压替换, 如式(10)所示:

$$\begin{cases} i_{s-d}(k+1) = \frac{T_s R_2}{L_d} i_{inv-d}(k) + \frac{T_s}{L_d} u_{c-d}(k) + \left[1 - \frac{T_s}{L_d}(R_s + R_2)\right] i_{s-d}(k) + \frac{T_s L_q \omega_e}{L_d} i_{s-q}(k) \\ i_{s-q}(k+1) = \frac{T_s R_2}{L_q} i_{inv-q}(k) + \frac{T_s}{L_q} u_{c-q}(k) - \frac{T_s L_d \omega_e}{L_q} i_{s-d}(k) + \left[1 - \frac{T_s}{L_q}(R_s + R_2)\right] i_{s-q}(k) - \frac{T_s \psi_f \omega_e}{L_q} \end{cases} \quad (10)$$

带 LC 滤波器的 PMSM 多步模型预测控制在当前 k 时刻预测未来多个时刻的逆变器电流、电容电压和电机电流。同样, 对于未来时刻的电机转速、电角度, 在采样频率较高的情况下, 预测模型认为转子转速近似保持不变, 转子角位置随转子角速度匀速变化, 逆变器不同开关状态输出的基本电压矢量在 d 、 q 轴上的分量随转子角位置而变化, 如式(11)~式(13)所示:

$$\omega_e(k+j) \approx \omega_e(k) \quad (11)$$

$$\theta_e(k+j) \approx \theta_e(k) + j \cdot \omega_e T_s \quad (12)$$

$$\begin{cases} u_{inv-d}(k+j) = u_\alpha \cos[\theta_e(k+j)] + u_\beta \sin[\theta_e(k+j)] \\ u_{inv-q}(k+j) = -u_\alpha \sin[\theta_e(k+j)] + u_\beta \cos[\theta_e(k+j)] \end{cases} \quad (13)$$

由此可得 LC 滤波器的 PMSM 多步模型预测控制的离散递推表达式模型为

$$\begin{cases} i_{inv-d}(k+j) = \frac{T_s}{L} u_{inv-d}(k+j-1) + \left[1 - \frac{T_s}{L}(R_1 + R_2)\right] i_{inv-d}(k+j-1) - \frac{T_s}{L} u_{c-d}(k+j-1) + \frac{T_s R_2}{L} i_{s-d}(k+j-1) \\ i_{inv-q}(k+j) = \frac{T_s}{L} u_{inv-q}(k+j-1) + \left[1 - \frac{T_s}{L}(R_1 + R_2)\right] i_{inv-q}(k+j-1) - \frac{T_s}{L} u_{c-q}(k+j-1) + \frac{T_s R_2}{L} i_{s-q}(k+j-1) \end{cases} \quad (14)$$

$$\begin{cases} u_{c-d}(k+j) = u_{c-d}(k+j-1) + \frac{T_s}{C} i_{inv-d}(k+j-1) - \frac{T_s}{C} i_{s-d}(k+j-1) \\ u_{c-q}(k+j) = u_{c-q}(k+j-1) + \frac{T_s}{C} i_{inv-q}(k+j-1) - \frac{T_s}{C} i_{s-q}(k+j-1) \end{cases} \quad (15)$$

$$\begin{cases} i_{s-d}(k+j) = \frac{T_s R_2}{L_d} i_{inv-d}(k+j-1) + \frac{T_s}{L_d} u_{c-d}(k+j-1) + \left[1 - \frac{T_s}{L_d}(R_s + R_2)\right] i_{s-d}(k+j-1) + \\ \frac{T_s L_q \omega_e(k)}{L_d} i_{s-q}(k+j-1) \\ i_{s-q}(k+j) = \frac{T_s R_2}{L_q} i_{inv-q}(k+j-1) + \frac{T_s}{L_q} u_{c-q}(k+j-1) - \frac{T_s L_d \omega_e(k)}{L_q} i_{s-d}(k+j-1) + \\ \left[1 - \frac{T_s}{L_q}(R_s + R_2)\right] i_{s-q}(k+j-1) - \frac{T_s \psi_f \omega_e(k)}{L_q} \end{cases} \quad (16)$$

式中: $i_{inv-d}(k+j)$ 、 $i_{inv-q}(k+j)$ 和 $i_{inv-d}(k+j-1)$ 、 $i_{inv-q}(k+j-1)$ 分别为 $k+j$ 时刻和 $k+j-1$ 时刻逆变器 d 、 q 轴电流; $u_{c-d}(k+j)$ 、 $u_{c-q}(k+j)$ 和 $u_{c-d}(k+j-1)$ 、 $u_{c-q}(k+j-1)$ 分别为 $k+j$ 时刻和 $k+j-1$ 时刻电容 d 、 q 轴电压; $i_{s-d}(k+j)$ 、 $i_{s-q}(k+j)$ 和 $i_{s-d}(k+j-1)$ 和 $i_{s-q}(k+j-1)$ 分别为 $k+j$ 时刻和 $k+j-1$ 时刻电机 d 、 q 轴电流; $\omega_e(k)$ 为预测过程中转子电角速度。

将 PMSM 多步电流预测模型转换为离散状

态空间模型, 其中将逆变器电流、电容电压和电机电流作为状态量, 如式 (17) 所示:

$$\mathbf{x} = [i_{inv-d} \quad i_{inv-q} \quad u_{c-d} \quad u_{c-q} \quad i_{s-d} \quad i_{s-q}]^T \quad (17)$$

由此, MPCC 的离散状态空间模型如式 (18) 所示:

$$\begin{cases} \mathbf{x}(k+1) = \mathbf{A}\mathbf{x}(k) + \mathbf{B}\mathbf{u}_{inv-dq}(k) + \mathbf{W} \\ \mathbf{y}(k) = \mathbf{x}(k) \end{cases} \quad (18)$$

$$\text{式中: } \mathbf{A} = \begin{bmatrix} 1 - \frac{T_s}{L}(R_1 + R_2) & 0 & -\frac{T_s}{L} & 0 & \frac{T_s R_2}{L} & 0 \\ 0 & 1 - \frac{T_s}{L}(R_1 + R_2) & 0 & -\frac{T_s}{L} & 0 & \frac{T_s R_2}{L} \\ \frac{T_s}{C} & 0 & 1 & 0 & -\frac{T_s}{C} & 0 \\ 0 & \frac{T_s}{C} & 0 & 1 & 0 & -\frac{T_s}{C} \\ \frac{T_s R_2}{L_d} & 0 & \frac{T_s}{L_d} & 0 & 1 - \frac{T_s}{L_d}(R_s + R_2) & \frac{T_s L_q \omega_e(k)}{L_d} \\ 0 & \frac{T_s R_2}{L_q} & 0 & \frac{T_s}{L_q} & -\frac{T_s L_d \omega_e(k)}{L_q} & 1 - \frac{T_s}{L_q}(R_s + R_2) \end{bmatrix}; \mathbf{B} = \begin{bmatrix} \frac{T_s}{L} & 0 \\ 0 & \frac{T_s}{L} \\ 0 & 0 \\ 0 & 0 \\ 0 & 0 \\ 0 & 0 \end{bmatrix};$$

$$\mathbf{W} = \begin{bmatrix} 0 \\ 0 \\ 0 \\ 0 \\ 0 \\ \frac{T_s \psi_f \omega_e(k)}{L_q} \end{bmatrix}。$$

利用线性变换将逆变器开关状态转换为逆变器输出 d 、 q 轴电压, 并将每一步预测的状态空间表达式合并, 可得 PMSM 三阶多步 MPCC 的离散状态空间线性模型如式 (19) 所示:

$$\mathbf{X} = \bar{\mathbf{A}}\mathbf{x}(k) + \bar{\mathbf{B}}\mathbf{U}(k) + \bar{\mathbf{W}} \quad (19)$$

1.2 成本函数

使用带 LC 滤波器的 PMSM 多步模型预测控制, 将逆变器输出电压作为输入的控制变量, 相比

于传统模型预测控制,增加了逆变器电流、电容电压作为状态量。因此,成本函数中,将逆变器电流、电容电压、电机电流及开关次数作为优化目标以发挥 LC 滤波器的作用,实现滤波的功能。

在多步模型预测控制中,每一步预测过程中的逆变器电流、电容电压及电机电流的参考值未知,同样,采用 k 时刻逆变器电流参考值 $i_{inv-d}^*(k)$ 和 $i_{inv-q}^*(k)$ 、电容电压参考值 $u_{c-d}^*(k)$ 和 $u_{c-q}^*(k)$ 及电机电流参考值 $i_{s-d}^*(k)$ 和 $i_{s-q}^*(k)$ 替代每一步预测过程中的参考值,如式 (20) 所示:

$$\begin{cases} i_{inv-d}^*(k) \approx i_{inv-d}^*(k+j), j=1,2,\dots,n \\ i_{inv-q}^*(k) \approx i_{inv-q}^*(k+j), j=1,2,\dots,n \\ u_{c-d}^*(k) \approx u_{c-d}^*(k+j), j=1,2,\dots,n \\ u_{c-q}^*(k) \approx u_{c-q}^*(k+j), j=1,2,\dots,n \\ i_{s-d}^*(k) \approx i_{s-d}^*(k+j), j=1,2,\dots,n \\ i_{s-q}^*(k) \approx i_{s-q}^*(k+j), j=1,2,\dots,n \end{cases} \quad (20)$$

为了调整成本函数中逆变器电流、电容电压、电机电流和开关次数之间的重要程度,对逆变器电流误差、电容电压误差、电机电流和开关次数增加权重系数。在离散递推表达式模型下,成本函数计算式为

$$\begin{aligned} g(n) = & \sum_{j=1}^n \lambda_1 \times \{ [i_{inv-d}(k+j) - i_{inv-d}^*(k)]^2 + \\ & [i_{inv-q}(k+j) - i_{inv-q}^*(k)]^2 \} + \\ & \sum_{j=1}^n \lambda_2 \times \{ [u_{c-d}(k+j) - u_{c-d}^*(k)]^2 + \\ & [u_{c-q}(k+j) - u_{c-q}^*(k)]^2 \} + \\ & \sum_{j=1}^n \lambda_3 \times \{ [i_{s-d}(k+j) - i_{s-d}^*(k)]^2 + \\ & [i_{s-q}(k+j) - i_{s-q}^*(k)]^2 \} + \\ & \sum_{j=1}^n \lambda \times Table[V(k+j-2), V(k+j-1)] \end{aligned} \quad (21)$$

式中: λ_1 为逆变器电流的权重系数; λ_2 为电容电压的权重系数; λ_3 为电机电流的权重系数; λ 为开关次数的权重系数; $Table$ 为开关切换次数表,根据上一时刻电压矢量 $V(k+i-2)$ 和当前时刻施加电压矢量 $V(k+i-1)$ 直接查表获取。

离散状态空间模型下,成本函数计算式为

$$g(n) = \| \bar{A}x(k) + \bar{B}U(k) + \bar{W} - \bar{Y}^*(k) \|_Q^2 + \lambda \| \bar{S}U(k) - \bar{E}U(k-1) \|_2^2 \quad (22)$$

式中: Q 为逆变器电流、电容电压和电机电流的权重系数矩阵; $Y^*(k)$ 为状态量参考值列向量,其包含逆变器电流、电容电压及电机电流参考值的列

$$\text{向量}; S = \begin{bmatrix} I_3 & 0 & 0 & \cdots & 0 \\ -I_3 & I_3 & 0 & \cdots & 0 \\ 0 & -I_3 & I_3 & \cdots & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & 0 & \cdots & I_3 \end{bmatrix}_{3n \times 3n}; E =$$

$$\begin{bmatrix} I_3 \\ 0 \\ 0 \\ \vdots \\ 0 \end{bmatrix}_{3n \times 3}$$

$U(k)$ 和 $U(k-1)$ 分别为当前时刻和上一时刻开关状态列向量。

1.3 最优电压矢量求解

PMSM 多步模型预测控制在每一步预测过程中有 8 个基本电压矢量作为备选控制量。因此, n 步模型预测控制共有 8^n 个电压矢量序列组合。遍历所有的电压矢量序列对 d, q 轴电流进行预测,依据预测值和参考值求得 8^n 个成本函数,通过对所有的成本函数寻优,得到最小成本函数对应的最优电压矢量序列,并将电压矢量序列的第一个电压矢量作为控制量作用于控制系统。

2 仿真分析

基于 Matlab/Simulink 建立带 LC 滤波器的 PMSM MPCC 模型,分别选取逆变器电流、电容电压和电机电流的权重系数为 10、0.5 和 500。PMSM 及 LC 滤波器参数如表 1 所示。

表 1 PMSM 及 LC 滤波器参数

Tab. 1 Parameters of PMSM and LC filter

参数名称	参数值
极对数 P_n	4
定子电阻 R_s/Ω	0.004
直轴电感 L_d/mH	0.94
交轴电感 L_q/mH	1.5
转子磁链 ψ_r/Wb	0.055
直流母线电压 U_{dc}/V	580
转动惯量 $J/(kg \cdot m^2)$	0.96
滤波电感 L_f/H	0.001
滤波电感等效阻抗 R_1/Ω	0.002
滤波电容 C_f/F	0.000 2
滤波电容等效阻抗 R_2/Ω	0.002

对于功率变换器驱动的电系统,降低电流总谐波失真(Total Harmonic Distortion, THD)和开关频率是重要的优化目标。因此,采用电流 THD 和平均开关频率作为谐波失真和开关损耗的评价指标。

电流 THD 定义为电流总谐波的均方根 I_H 与电流基波信号的均方根 I_F 之比,如式(23)所示:

$$I_{\text{THD}} = \frac{I_H}{I_F} = \frac{\sqrt{I_2^2 + I_3^2 + \dots + I_n^2}}{I_F} \quad (23)$$

本文采用 A、B 和 C 三相电流 THD 的平均值来计算电流 THD,用以表征驱动系统的谐波失真,如式(24)所示:

$$I_{\text{THD}} = \frac{I_{\text{A-THD}} + I_{\text{B-THD}} + I_{\text{C-THD}}}{3} \quad (24)$$

平均开关频率用以表征驱动系统的开关损耗,如式(25)所示:

$$f_{\text{sw}} = \frac{N_{\text{switch}}}{6t} \quad (25)$$

式中: N_{switch} 为系统运行过程中逆变器开关总次数; t 为系统运行时长。

本文将 d 、 q 轴电流脉动均方根误差(Root Mean Square Error, RMSE)作为衡量电流脉动的评价指标,将转矩脉动 RMSE 作为衡量转矩脉动的评价指标,分别如式(26)~式(28)所示:

$$i_{d\text{-RMSE}} = \sqrt{\frac{1}{N} \sum_{n=1}^N (i_{dn} - \hat{i}_d)^2} \quad (26)$$

$$i_{q\text{-RMSE}} = \sqrt{\frac{1}{N} \sum_{n=1}^N (i_{qn} - \hat{i}_q)^2} \quad (27)$$

$$T_e\text{-RMSE} = \sqrt{\frac{1}{N} \sum_{n=1}^N (T_{en} - \hat{T}_e)^2} \quad (28)$$

式中: N 为数据总个数, \hat{i}_d 和 \hat{i}_q 分别为定子 d 、 q 轴电流参考值; \hat{T}_e 为转矩参考值。

将开关频率设置在 5.0 kHz 附近,其中,传统不带 LC 滤波器的一阶 MPCC 权重系数 $\lambda = 58$,开关频率 $f_{\text{sw}} = 4.9773$ kHz;带滤波器的三阶 MPCC 权重系数 $\lambda = 700$,开关频率 $f_{\text{sw}} = 5.0023$ kHz。设置预测步长为五步,一阶 MPCC 和三阶 MPCC 仿真波形如图 3~图 10 所示。

由仿真结果可知,相较于一阶五步 MPCC,三阶五步 MPCC 的控制性能得到有效提升。相同开关频率下,转矩和电流脉动明显降低,电流 THD

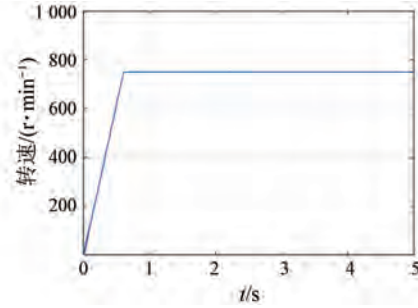


图 3 一阶 MPCC 转速

Fig. 3 Speed of first-order MPCC

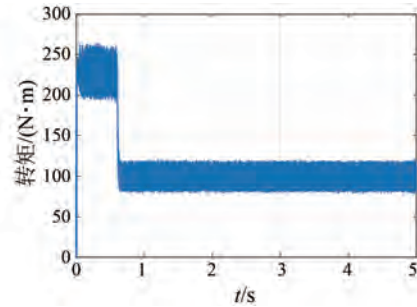


图 4 一阶 MPCC 转矩

Fig. 4 Torque of first-order MPCC

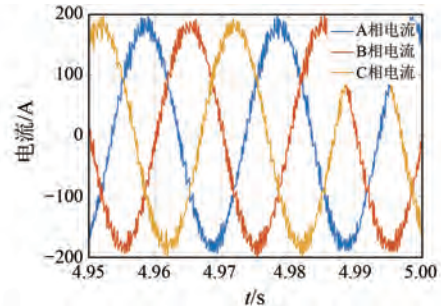


图 5 一阶 MPCC 三相定子电流

Fig. 5 Three-phase stator current of first-order MPCC

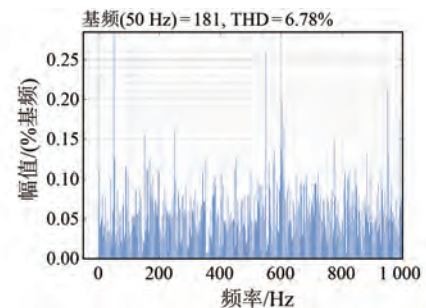


图 6 一阶 MPCC 的 A 相定子电流 THD

Fig. 6 THD of A-phase stator current of first-order MPCC

更低。

在基准转速 750 r/min 和额定转矩 100 N·m

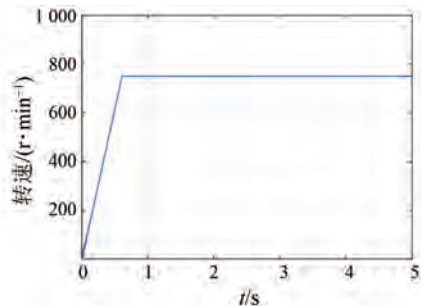


图 7 三阶 MPCC 转速

Fig. 7 Speed of third-order MPCC

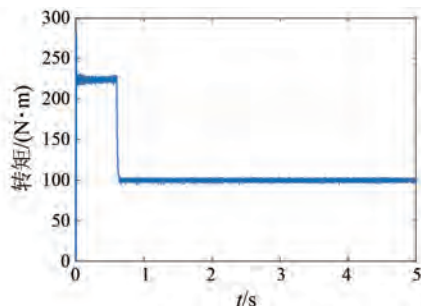


图 8 三阶 MPCC 转矩

Fig. 8 Torque of third-order MPCC

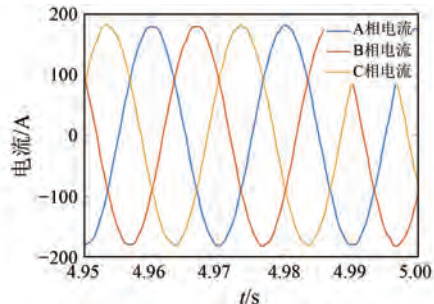


图 9 三阶 MPCC 三相定子电流

Fig. 9 Three-phase stator current of third-order MPCC

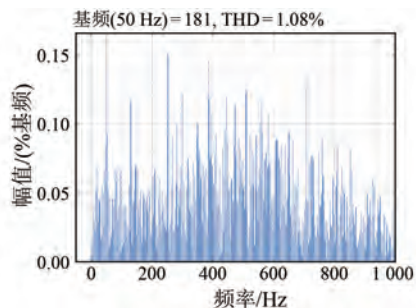


图 10 三阶 MPCC 的 A 相定子电流 THD
 Fig. 10 THD of A-phase stator current of third-order MPCC

的条件下进行仿真。通过调整开关次数权重系数得到不同开关频率下的 LC 滤波器 MPCC 性能。

以平均开关频率 f_{sw} 作为横坐标, 电流谐波含量 I_{THD} 作为纵坐标, 得到 $I_{THD}-f_{sw}$ 散点图, 如图 11~图 13 所示。

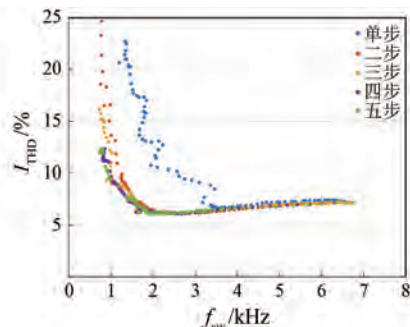


图 11 一阶 MPCC $I_{THD}-f_{sw}$ 散点图

Fig. 11 Scatterplot of $I_{THD}-f_{sw}$ of first-order MPCC

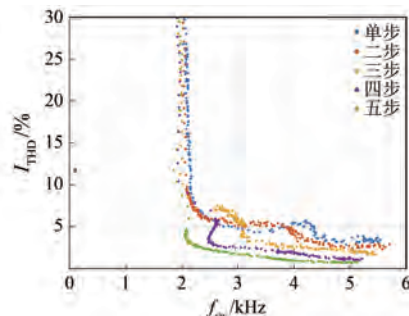


图 12 三阶 MPCC $I_{THD}-f_{sw}$ 散点图

Fig. 12 Scatterplot of $I_{THD}-f_{sw}$ of third-order MPCC

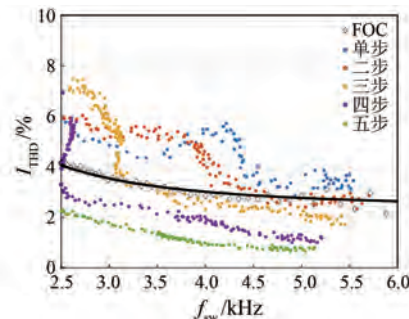


图 13 三阶 MPCC 和 FOC $I_{THD}-f_{sw}$ 散点图

Fig. 13 Scatterplot of $I_{THD}-f_{sw}$ of third-order MPCC and FOC

由仿真结果可知: 对于一阶 MPCC, 当开关频率大于 3.5 kHz 时, 一步至五步的控制性能基本相同; 当开关频率低于 3.5 kHz 时, 多步预测控制性能优于单步控制, 但两步至五步的控制性能基本相同。对于三阶 MPCC, 随着预测步数的增加, 控制性能逐步提升。相同开关频率和预测步数下, 三阶 MPCC 的控制性能优于一阶 MPCC。当

开关频率大于 2.5 kHz 时,相同开关频率下,四步至五步的三阶 MPCC 控制性能优于 FOC。

将开关频率设置在 5.0 kHz 附近,对比一阶 MPCC、三阶 MPCC 和 FOC 的控制性能,结果如表 2 和图 14~17 所示。

表 2 一阶 MPCC、三阶 MPCC 和 FOC 控制性能对比

Tab. 2 Comparison of control performance of first-order MPCC, third-order MPCC, and FOC

控制系统名称	步长	调整参数	$f_{sw}/$ kHz	$I_{THD}/$ %	$i_{d-RMSE}/$ A	$i_{q-RMSE}/$ A	$T_e-RMSE/$ (N·m)
一阶 MPC	1	$\lambda = 54$	5.00	7.23	10.44	8.71	7.96
	2	$\lambda = 58$	4.98	6.91	10.26	8.09	7.54
	3	$\lambda = 85$	5.01	7.00	10.51	8.36	7.72
	4	$\lambda = 70$	5.00	6.91	10.61	8.10	7.52
	5	$\lambda = 58$	4.98	6.8	10.73	7.56	7.23
三阶 MPC	1	$\lambda = 275$	4.90	4.19	6.90	8.91	5.26
	2	$\lambda = 160$	5.17	3.63	6.14	8.34	4.87
	3	$\lambda = 150$	5.00	3.03	3.62	6.68	3.65
FOC	4	$\lambda = 500$	5.00	1.51	2.25	4.04	1.85
	5	$\lambda = 700$	5.00	1.08	2.33	2.82	1.16
FOC	无	$T_s = 4 \times 10^{-4}$	5.00	2.86	3.73	1.45	2.68

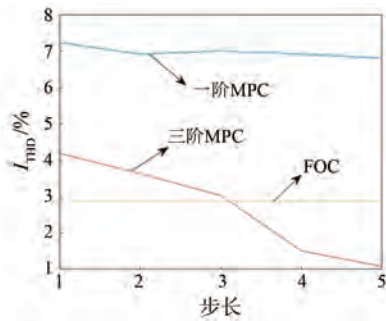


图 14 一阶 MPCC、三阶 MPCC 及 FOC 的电流 THD

Fig. 14 THD of current of first-order MPCC, third-order MPCC and FOC

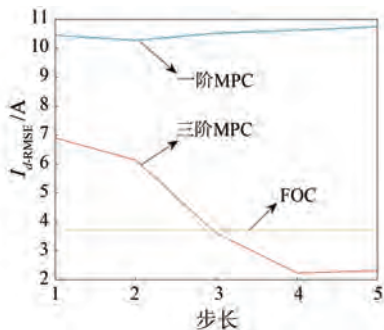


图 15 一阶 MPCC、三阶 MPCC 及 FOC 的 d 轴电流脉动 RMSE

Fig. 15 RMSE of d-axis stator current ripple of first-order MPCC, third-order MPCC and FOC

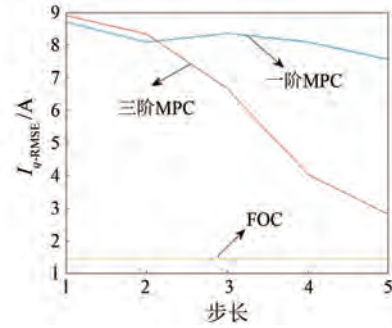


图 16 一阶 MPCC、三阶 MPCC 及 FOC 的 q 轴电流脉动 RMSE

Fig. 16 RMSE of q-axis stator current ripple of first-order MPCC, third-order MPCC and FOC

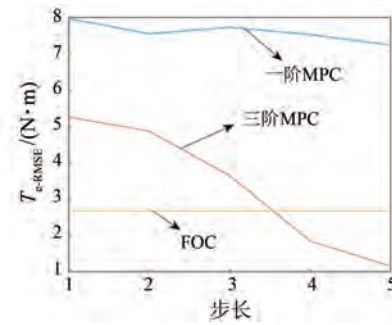


图 17 一阶 MPCC、三阶 MPCC 及 FOC 的转矩脉动 RMSE

Fig. 17 RMSE of torque ripple of first-order MPCC, third-order MPCC and FOC

由仿真结果可知,开关频率在 5.0 kHz 附近,随着预测步数的增加,一阶 MPCC 的控制性能基本不变。随着预测步长的增加,三阶 MPCC 的控制性能逐步提升,电流 THD、d、q 轴电流脉动 RMSE 和转矩脉动 RMSE 逐步减小。相同预测步数下,三阶 MPCC 的控制性能明显优于一阶 MPCC。当预测步数超过三步,三阶 MPCC 具有更小的电流谐波和转矩脉动,控制效果优于 FOC。

3 结语

(1) 当开关频率大于 3.5 kHz 时,一阶 MPCC 一步至五步的控制性能基本相同;当开关频率低于 3.5 kHz 时,一阶 MPCC 多步控制性能优于单步控制,但两步至五步的控制性能基本相当。

(2) 随着预测步数的增加,三阶 MPCC 的控制性能逐步提升;相同开关频率和预测步数下,三阶 MPCC 的控制性能优于一阶 MPCC。

(3) 当开关频率大于 2.5 kHz 时,相同开关

频率下,步长为四步、五步的三阶 MPCC 比 FOC 的控制性能更优。

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收稿日期:2024-04-18

收到修改稿日期:2024-07-26

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Multi-Step Model Predictive Current Control of Permanent Magnet Synchronous Motor with LC Filter

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Key words: permanent magnet synchronous motor; model predictive current control; multi-step prediction; LC filter

Model predictive current control (MPCC) for permanent magnet synchronous motors (PMSMs) has garnered significant attention due to its simple structure, flexible control, and no need for pulse width modulation. However, because of the limited number of candidate voltage vectors, MPCC exhibits relatively large current and torque ripples, resulting in inferior control performance compared to field oriented control (FOC). To improve control performance, an LC filter was introduced into the

PMSM system, and a third-order MPCC was developed. Predictive control was applied to the inverter current error, capacitor voltage error, motor current error, and switching frequency. The control performance of the third-order MPCC was compared with that of the first-order MPCC and FOC.

At an average switching frequency of around 5 kHz, the control performance of the first-order MPCC, third-order MPCC, and FOC was shown in Tab.1.

Tab.1 Control performance of first-order MPCC, third-order MPCC and FOC

Control system	Predictive step	Weighting factor	f_{sw}/kHz	$I_{THD}/\%$	i_{d-RMSE}/A	i_{q-RMSE}/A	$T_{e-RMSE}/(\text{N}\cdot\text{m})$
First-order MPCC	1	$\lambda = 54$	5.00	7.23	10.44	8.71	7.96
	2	$\lambda = 58$	4.98	6.91	10.26	8.09	7.54
	3	$\lambda = 85$	5.01	7.00	10.51	8.36	7.72
	4	$\lambda = 70$	5.00	6.91	10.61	8.10	7.52
	5	$\lambda = 58$	4.98	6.8	10.73	7.56	7.23
Third-order MPCC	1	$\lambda = 275$	4.90	4.19	6.90	8.91	5.26
	2	$\lambda = 160$	5.17	3.63	6.14	8.34	4.87
	3	$\lambda = 150$	5.00	3.03	3.62	6.68	3.65
	4	$\lambda = 500$	5.00	1.51	2.25	4.04	1.85
	5	$\lambda = 700$	5.00	1.08	2.33	2.82	1.16
FOC	/	$T_s = 4 \times 10^{-4}$	5.00	2.86	3.73	1.45	2.68

The conclusions are as follows:

1. For the first-order MPCC, when the switching frequency exceeds 3.5 kHz, the control performance from one to five prediction steps is roughly equivalent. When the switching frequency is below 3.5 kHz, multi-step prediction outperforms single-step prediction, though the performance from two to five steps is approximately the same.

2. For the third-order MPCC, the control

performance improves as the number of prediction steps increases.

3. At the same switching frequency and number of prediction steps, the control performance of the third-order MPCC surpasses that of the first-order MPCC. When the switching frequency exceeds 2.5 kHz, the four-step and five-step third-order MPCC outperforms FOC at the same switching frequency.